TITLE OF THE INVENTION

ACTIVE RESISTIVE SUMMER FOR A TRANSFORMER HYBRID

CROSS-REFERENCE TO RELATED CASES

[0001] This application is a continuation-in-part of and claims benefit of priority of co-pending U.S. patent application Serial No. 09/629,092 entitled "ACTIVE RESISTIVE SUMMER FOR A TRANSFORMER HYBRID," by the same inventors, filed July 31, 2000, the disclosure of which is hereby incorporated by reference. This application is also related to U.S. patent application Serial No. [not yet assigned] entitled "APPARATUS AND METHOD FOR CONVERTING SINGLE-ENDED SIGNALS TO A DIFFERENTIAL SIGNAL, AND TRANSCEIVER EMPLOYING SAME," by Pierte Roo, filing date [concurrent herewith], the disclosure of which is hereby incorporated by reference.

BACKGROUND OF THE INVENTION

Field of the Invention

[0002] The present invention relates generally to transmitting and receiving electrical signals through communication channels, such as a gigabit channel. In particular, the present invention relates to a transmit canceller that removes transmit signals from receive signals in such communication channels.

Background and Related Art

[0003] A gigabit channel is a communications channel with a total data throughput of one gigabit per second. A gigabit channel typically includes four (4) unshielded twisted pairs (hereinafter "UTP") of cables (e.g., category 5 cables) to achieve this data rate. IEEE Standard 802.3ab, herein incorporated by reference, specifies the physical layer parameters for a 1000BASE-T channel (e.g., a gigabit channel).

[0004] As will be appreciated by those skilled in the art, a UTP becomes a transmission line when transmitting high frequency signals. A transmission line can be modeled as a network of inductors, capacitors and resistors, as shown in Figure 1. With reference to Figure 1, G is normally zero and R(ω ω

$$R(\omega) = k_R(1+j)\sqrt{\omega}, \qquad (1)$$

where k_R is a function of the conductor diameter, permeability, and conductivity. The characteristic impedance of the line is defined by:

$$Z_0 = \sqrt{\frac{R(\omega)+j\omega L}{G+j\omega C}}, \tag{2}$$

and at high frequencies, Z_0 becomes approximately $\sqrt{L/C}$ or approximately 100 ohms in a typical configuration. When properly terminated, a UTP of length d has a transfer function H that is a function of both length (d) and frequency (ω):

$$H(d,\omega) = e^{d\gamma(\omega)}, \tag{3}$$

where

$$\gamma\omega = \sqrt{(R(\omega)+j\omega L)(G+j\omega C)}, \tag{4}$$

and substituting Equations 1 and 4 into Equation 3, and simplifying, approximately yields:

$$H(d,\omega) \approx \exp\left\{d\left[\frac{k_\kappa}{2}\sqrt{\frac{\omega L}{C}}+j\left(\omega\sqrt{LC}+\frac{k_\kappa}{2}\sqrt{\frac{\omega L}{C}}\right)\right]\right\}. \tag{5}$$

Equation 5 shows that attenuation and delay are a function of the cable length d.

[0005] A transmission path for a UTP typically includes a twisted pair of cables that are coupled to transformers at both a near and far end, as shown in Figure 2. A transceiver at each end of the transmission path transmits

and receives via the same twisted pair. A cable typically includes two patch cords totaling less than 10m, and a main section of 100m or even longer. The transmitters shown in Figure 2 are modeled as current sources. The near end current source supplies a current I_{tx}. The near end transmit voltage (e.g., $I_{tx}R_{tx}$) is detected and measured across resistor R_{tx}. A receive signal V_{rcv} (e.g., a signal transmitted from the far-end transceiver) is also detected and measured across resistor R_{tx}. Hence, V_{tx} includes both transmit ($I_{tx}R_{tx}$) and receive (V_{rcv}) signals. Accordingly, the signal V_{rcv} (e.g., the signal from Transceiver B) received at Transceiver A can be obtained by taking the difference between the transmit voltage and the measured voltage V_{tx}, as follows:

$$V_{rcv} = V_{tx} - I_{tx}R_{tx}. \tag{6}$$

[0006] Conventional solutions for removing transmit signals from receive signals often employ known transconductor ("Gm") summing stages or other current based methods. As will be appreciated, these methods often introduce signal distortion into the receive signal. Also, some transconductors have a limited signal dynamic range. Accordingly, conventional methods are often inadequate for applications requiring signal recovery. Additionally, known

summing circuits, such as weighted summers using operational amplifiers, have not heretofore been modified to accommodate the intricacies associated with canceling transmit signals or regulating baseline wander (described below). A known weighted summer is discussed in Chapter 2 of "Microelectronic Circuits, Third Edition," by A.S. Sedra and K.C. Smith, 1991, incorporated herein by reference.

[0007] As will be appreciated by those skilled in the art, the receive signal V_{rcv} typically contains additional components, due to baseline wander, echoes and crosstalk, for example.

[0008] Baseline wander is preferably corrected for when transmitting and receiving signals over transmission lines. Removing DC components from a receive signal using transformer coupling can cause baseline wander. As will be appreciated by those skilled in the art, baseline wander represents a deviation from an initial DC potential of a signal.

[0009] "Echoes" typically represent a residual transmit signal caused by reflections that appear in the receive signal. Echoes can cause undue interference depending on

the size of the reflection.

[00010] Capacitive coupling between the channels, as shown in Figure 3, causes crosstalk. Four channels TX1-TX4 are shown in Figure 3. The capacitive coupling between TX1 and each of TX2, TX3 and TX4 are modeled by capacitors C_{1-2}, C_{1-3}, C_{1-4}, respectively. The capacitive coupling forms a high-pass filter between channels and therefore crosstalk contains mostly high frequency components. As will be appreciated by those skilled in the art, normally only the near-end crosstalk (NEXT) needs to be considered, since crosstalk is usually small and the transmission line provides further attenuation of the far-end crosstalk (FEXT).

[0010] Accordingly, there are many signal-to-noise problems to be solved in the art. Hence, an efficient transmission canceller is needed to remove a transmit signal from a receive signal without introducing excess signal distortion. An electrical circuit is also needed to subtract a transmit signal from a receive signal. There is a further need of an electrical circuit to correct baseline wander.

SUMMARY OF THE INVENTION

[0011] The present invention relates to a transmit signal canceller for use in a transformer hybrid. Such a hybrid includes a junction for transmitting and receiving signals. In the present invention, an active resistive summer can be used to cancel a transmit signal from a receive signal.

[0012] According to the invention, an electrical circuit in a communications channel is provided. The electrical circuit includes an active resistive summer having: (i) an input for a composite signal, the composite signal including a transmission signal component and a receive signal component, (ii) an input for a replica transmission signal, and (iii) an output for a receive signal which includes the composite signal minus the replica signal.

[0013] According to an another aspect of the present invention, a transmit signal canceller in a communication channel is provided. The channel includes a first transceiver for transmitting and receiving signals and a replica transmitter for generating a replica transmission signal input. A composite signal at a near end includes a transmission signal of the first transceiver and a received

signal of a second transceiver. The transmit canceller includes: (i) an operational amplifier having a positive input terminal, a negative input terminal, and an output terminal; (ii) a feedback element in communication with the negative input terminal and the output terminal; (iii) a first input resistor in communication with the negative input terminal and the measured signal input; (iv) a second input resistor in communication with the negative input terminal and the replica signal input; and (v) a predetermined voltage source in communication with the positive terminal of the operational amplifier. The receive signal is an output at the output terminal of the operational amplifier.

[0014] According to still another aspect of the present invention, a communication system including a first transmission channel with a first end and a second end is provided. The first end couples to a first transformer and the second end couples to a second transformer. A first transceiver transmits and receives signals via the first transformer and a second transceiver transmits and receives signals via the second transformer. A first signal is supplied at the near end. The first signal includes a transmission signal component of the first transceiver and

a receive signal component of the second transceiver. The communications system includes: (i) a replica transmitter that generates a replica of the transmission signal component of the first transceiver; (ii) a filter to filter the replica signal; (iii) an active resistive summer receiving the first signal, and the filtered replica signal as inputs to reduce the transmission signal component at an output of the active resistive summer.

[0015] According to still another aspect of the present invention, a method of correcting baseline wander in a receive signal in a communications channel having a near and far end is provided. The channel includes a first transceiver at the near end and a second transceiver at the far end, each to transmit and receive signals. The method includes the steps of: (i) providing a composite signal, the composite signal including a transmission signal of the first transceiver and a receive signal of the second transceiver; (ii) generating a replica of the transmission signal; (iii) subtracting the replica signal from the composite signal through an active resistive summer; and (iv) providing a baseline correction current into the active resistive summer.

[0016] According to still another aspect of the present invention, an electrical circuit in a communications system is provided. A composite signal including a transmission signal component and a receive signal component, a replica transmission signal and a common-mode shift current are provided. Further circuitry is provided to control the magnitude of the common-mode shift current so that the magnitude of the composite signal does not exceed a predetermined value of an operating parameter of the electrical circuit.

[0017] In still another aspect of the present invention, an electrical circuit in a communications system is provided. An active resistive summing circuit produces a receive signal as a difference between a composite signal and a replica transmission signal, the composite signal comprising a transmission signal component and a receive signal component. Further circuitry is provided which controls the magnitude of the composite signal.

[0018] In still another aspect of the present invention, another electrical circuit in a communications system is provided. An active resistive summer is provided that receives a composite signal that includes a transmission

signal component and a receive signal component, a replica transmission signal, and a common-mode shift current signal. The active resistive summer provides an output which is a receive signal that comprises the composite signal minus the replica signal. Further circuitry is provided which controls the magnitude of the common-mode shift current to thereby control the magnitude of the composite signal.

[0019] These and other objects, features, and advantages of the present invention will be apparent from the following description of the preferred embodiments of the present invention.

BRIEF DESCRIPTION OF THE DRAWINGS

[0020] The details of the present invention will be more readily understood from a detailed description of the preferred embodiments taken in conjunction with the following figures.

[0021] Figure 1 is a circuit diagram illustrating a transmission line model.

[0022] Figure 2 is a circuit diagram illustrating a transmission path across a twisted pair of cables, the cables being coupled to transformers at each end.

[0023] Figure 3 is a diagram-illustrating crosstalk between channels in a gigabit channel.

[0024] Figure 4 is a block diagram illustrating a system overview of a communications channel.

[0025] Figure 5 is a circuit diagram illustrating a transmitter.

[0026] Figure 6 is a graph illustrating a transmit signal.

[0027] Figure 7 is a graph illustrating a composite signal with echoes.

[0028] Figure 8 is a circuit diagram illustrating a replica transmitter.

[0029] Figure 9 is a graph illustrating a receive signal.

[0030] Figure 10 is block diagram illustrating a low-pass filter.

[0031] Figure 11 is a circuit diagram illustrating an active resistive summer.

[0032] Figure 12 is a circuit diagram illustrating an error detection circuit.

[0033] Figure 13 is a circuit diagram illustrating a low-pass filter.

[0034] Figure 14 is a circuit diagram illustrating a conventional voltage controlled current source.

[0035] Figure 15 is a circuit diagram illustrating one exemplary embodiment of a common-mode feedback circuit coupled with a transmitter.

[0036] Figure 16 is a circuit diagram illustrating one exemplary implementation of an operational amplifier utilized in the common-mode feedback circuit.

[0037] Figure 17 is a circuit diagram illustrating one exemplary embodiment of a common-mode shift current control circuit coupled with a transmitter.

[0038] Figure 18 is a circuit diagram illustrating one exemplary embodiment of a common-mode shift control circuit coupled with a transmitter.

DETAILED DESCRIPTION OF THE PRESENTLY PREFERRED EMBODIMENTS

[0039] The preferred embodiments will be described with respect to a gigabit channel, as used, for example, in an Ethernet network; and to electrical circuits associated with separating transmit and receive signals in such a gigabit channel. The preferred embodiments will also be described with respect to baseline wander correction in such a gigabit channel. However, as will be appreciated by those skilled in the art, the present invention is also applicable to other transmission channels, and to other electrical circuits having applications requiring cancellation of transmit signals, for example.

[0040] Figure 4 is a block diagram illustrating principle components for one of the four channels in a

preferred gigabit channel configuration for use in an Ethernet network. As illustrated in Figure 4, a vertical dashed line divides analog and digital processing components. The analog components preferably include a transmitter ("XMTR") 1, replica transmitter ("Replica XMTR") 2, transmit canceller 3, baseline correction module 4, low pass filter ("LPF") 5, analog-to-digital converter ("ADC") 6, and phase-lock loop ("PLL") 7. A known PLL can be used with the present invention.

[0041] Digital processing components preferably include a transmitter encoder 10, echo module 11, NEXT cancellers 12-14 to assist in removing echoes, synchronization module 15, FIR (Finite Impulse Response) equalizer 16 and a DFE (Decision Feedback Equalizer) 17 to equalize a receive signal, and a Viterbi module 18. The digital processing components also include baseline correction modules 19 and 20 to correct residual baseline wander. A timing recovery module 21, an error correction detector 22 (described in further detail below), and summing junction 23 are also shown. The individual digital components designated by blocks in Figure 4 are all well known in the communication arts, and their specific construction and operation are not critical to the operation or best mode for carrying out the

present invention.

[0042] The analog "front-end" components shown in Figure 4 will now be described in even further detail. The front-end analog components are preferably designed and constructed via customized integrated circuits. However, as will be appreciated by those skilled in the art, the inventive circuits and corresponding configuration could also be realized using discrete components as well.

[0043] As illustrated in Figure 5, transmitter 1 preferably includes a current-source I_{tx} that generates a transmit signal over a resistor R_{tx}. An appropriate value for resistor R_{tx} can be selected to match the line impedance, for example. In one preferred embodiment, a resistor center tap is set to 2.5 volts so the transmitter 1 effectively sees a differential impedance of 25 ohms. Preferred performance specifications for the transmitter 1 are further detailed in Table 1, below.

[0044] An impulse transmit signal can be generated from a unit square pulse of 1T width filtered by a one-pole, low-pass filter (not shown) with a cutoff frequency between 85MHz and 125MHz. Slew-rate control can also be used to

limit the rise and fall times and thus reduce the high frequency components of a transmit signal. Of course, any transmit signal preferably fits into the transmit template provided by the IEEE 802.3ab Standard. An ideal transmit pulse is shown in Figure 6.

[0045] A measured voltage V_{tx} across R_{tx} (Figure 5) is shown in Figure 7. The measured signal V_{tx} contains interference caused by line reflections (e.g., echoes). The reflections are caused by impedance discontinuity due to impedance mismatch between different cables. For example, a large reflection pulse at 60ns as shown in Figure 7 corresponds to a reflection from the impedance discontinuity at an adapter connecting a 5m patch cord to a 100m cable. The magnitude of the echoes can be significant when compared to the magnitude of the receive signal at a long line length, and therefore, echo cancellation, as provided by the NEXT cancellers 12-14 shown in Figure 4, is employed.

[0046] A receive signal V_{rcv} (e.g., a signal received from a far-end transceiver) is also measured across resistor R_{tx}, as shown in Figure 5. Accordingly, the near end transmit signal ($I_{tx}R_{tx}$) is preferably canceled or

reduced from the composite signal V_{tx} in order to effectively recover the far-end received signal V_{rcv}. This type of active cancellation can be accomplished with a replica transmit signal V_{txr}. Accordingly, a replica transmitter 2' (to be described below) is provided to generate a signal V_{txr} to be subtracted from the measured signal V_{tx}, thus, effectively reducing the transmit signal $(I_{tx}R_{tx})$.

[0047] A receive signal x(t) transmitted with pulse amplitude modulation ("PAM") is define by:

$$x(t) = \sum_{n=1}^{\infty} a_n p(t - nT), \qquad (7)$$

where a_n is the transmit symbols and $p(t)$ is the channel pulse derived by convoluting an impulse transmit pulse with a channel response defined by Equation 5. The receive signal for a 100m cable is heavily attenuated by the transmission line and the pulse width is dispersed, as shown in Figure 9. A 100m UTP delays the signal by about 550ns. Signal equalization preferably uses high frequency boosting via the FIR 16 to remove precursor intersymbol interference ("ISI") and to insert a zero crossing for timing recovery 21. The DFE 17 is used to remove

postcursor ISI.

[0048] The receive signal's elongated tail results from transformer coupling (e.g., a high-pass filter) with a time constant (e.g., L/R) typically on the order of micro-seconds. Since the receive signal contains little or no average DC energy, the negative tail has the same amount of energy as the positive pulse. In this regard, the signal's area integral is zero. In a typical example, a tail can last over 10μs with a magnitude of no more than 0.5mV. The long tail causes any DC bias to drift back toward zero, which can lead to baseline wander. As will be appreciated, this response time is too long to be practically removed by a digital equalizer, but the response is slow enough to be cancelled using a slow integrator, for example. The baseline wander canceller 4 is preferably decision directed to minimize the error defined by the difference between the equalized value and it's sliced value, as discussed below.

[0049] As illustrated in Figure 8, the replica transmitter 2 includes a current source I_{txr}. I_{txr} is coupled to a voltage V through resistors R, as shown in Figure 8. In a preferred embodiment, R is 100 ohms and V is about 2.5 volts. The replica signal V_{txr} is preferably

filtered through a known low-pass filter to obtain a low-pass replica signal ("V_{txrl}"), as shown in Figure 10. Replica signal V_{txr} can also be inverted in a known manner to produce $-V_{txr}$. The preferred performance specifications for the transmitter 1 and replica transmitter 2 are shown in Table 1.

Table 1: Transmitter and Replica Performance Specifications

Parameters	Specifications
Transmit Current	+/- 40mA
Replica Transmit Current	¼ of transmit current
Number of levels	16 (not including 0)
Number of sub-units	8 (sequentially delayed)
Transmit Profile	[1 1 2 2 1 1], w/ ~1ns delay
Replica Transmit Profile	[1 1 3 3], w/ ~1ns delay
R_{tx}	100Ω

[0050] A transmit signal canceller 4 is illustrated in Figure 11. The transmit canceller 4 removes the transmission signal ($I_{tx}R_{tx}$) from the measured (or detected) transmit V_{tx} signal. In particular, the transmit canceller includes an active resistive summer that provides a large input dynamic range and stable linearity characteristics, while removing (e.g., reducing or canceling) the unwanted transmit signal component.

[0051] As illustrated in Figure 11, the active summer includes an operational amplifier ("op-amp") with inverting

feedback. The op-amp is preferably constructed using integrated circuits in a known manner. The summer receives V_{txrl}, V_{tx}, $-V_{txr}$, I_{cms}, and I_{bl} as input signals. I_{bl} is a baseline wander control current, and I_{cms} is a common-mode shift current, each as further discussed below.

[0052] As will be appreciated by those skilled in the art, a transformer typically has high-pass characteristics. Accordingly, replica signal $-V_{txr}$ is combined (e.g., subtracted via the active resistive summer) with the low pass replica signal V_{txrl} to produce a high-pass replica signal. As an alternative configuration, V_{txr} could be filtered through a known high-pass filter prior to the transmit canceller 3 stage.

[0053] Returning to Figure 11, receive signal V_{rcv} is determined from the following relationships.

[0054] Let:

Vi = voltage for the op-amp's positive terminal;

$V_1 = V_{txrl}$;

$V_2 = V_{tx}$;

$-V_3 = -V_{txr}$;

$i_4 = I_{cms}$; and

$i_5 = I_{bl}$.

Then:

$i_1 + i_2 - i_3 - i_4 - i_5 = i_0$; and

$$\frac{V_1 - V_i}{R_1} = i_1; \quad \frac{V_2 - V_i}{R_1} = i_2; \quad \frac{V_i - -V_3}{R_1} = i_3; \quad \frac{V_i - V_{rcv}}{R_f} = i_0.$$

$$\frac{V_1 - V_i}{R_1} + \frac{V_2 - V_i}{R_1} - \frac{V_i + V_3}{R_1} - i_4 - i_5 = \frac{V_i - V_{rcv}}{R_f}$$

$$\frac{V_1 + V_2 - V_3 - 3V_i}{R_1} - i_4 - i_5 = \frac{V_i - V_{rcv}}{R_f}$$

$$\frac{R_f}{R_1}(V_1 + V_2 - V_3 - 3V_i) - i_4\ R_f - i_5 \cdot R_f = V_i - V_{rcv}$$

$$\frac{R_f}{R_1}(V_1 + V_2 - V_3 - 3v_i) - R_f i_4 - R_f i_5 - V_i = -V_{rcv}$$

$$V_{rcv} = V_i - \frac{R_f}{R_1}(V_1 + V_2 - V_3 - 3V_i) + R_f(i_4 + i_5) \qquad (8)$$

[0055] Substituting the input signals for their placeholders yields the definition for V_{rcv}, as follows:

$$V_{rcv} = V_i - \frac{R_F}{R_1}(V_{txrl} + V_{tx} - V_{txr} - 3V_i) + R_f(I_{cms} + I_{bl}). \qquad (9)$$

The gain is preferably set between 0.75 and 1 (e.g., R_F/R_1 equals 0.75 to 1). For a small signal analysis, Vi can be set to zero (0). Also, as will be appreciated by those skilled in the art, in a fully differential circuit, Vi effectively drops out of the equations since $V_{rcv} = V_{rcv}^{(+)} - V_{rcv}^{(-)}$. As discussed, V_{txrl} and $-V_{txr}$ are combined through the

active summer to provide a high-pass replica signal ("V_{txrh}"). The receive signal V_{rcv} can then be recovered as shown by Equation 9.

[0056] Preferred transmit canceller specifications are detailed in Table 2, below.

Table 2: Transmit Canceller Performance Specifications

Parameters	Specifications
Input Dynamic Range	+/-2.5V(diff.) for transmit signal
Output Dynamic Range	+/-1V(diff.)
Input impedance	High, ~10k.
Output impedance	Low
Cutoff frequency	Greater than 31.5 Mhz
DC Gain	0.85 -dependent on the LPF 5 and ADC 6 characteristics (Figure 4)
Power	25 mw, including LPF 5 (Figure 4)
R_f	8.5 KΩ; or 7.5KΩ for increased attenuation
Vi	2.0 volts
R_1	10KΩ

[0057] A known current mode circuit, e.g., a voltage controlled current source (VCCS) as shown in Figure 14, with feedback preferably sets the summer input current-mode voltage (V_{cm}). Of course, other known current mode circuits could be employed with the present invention. This current-mode circuit shifts the common-mode of both the transmit and replica transmit signals. The input to the op

amp (V_{aip}, V_{ain}) is compared against the desired op amp output common-mode voltage (V_d):

$$V_d = (V_{aip} - V_{cm}) + (V_{ain} - V_{cm}). \qquad (10)$$

Then, the common-mode shift current can be determined from:

$$I_{cms} = V_d g_m + I_0, \qquad (11)$$

where g_m is a transconductance and I_o is an offset current. An appropriate transconductance and offset current can be selected by setting $V_{cm} = I_{cms}R_F = V_d g_m R_F + I_0 R_F$, to ensure a proper common-mode voltage seen by the op amp inputs. In this manner, the common mode shift current I_{cms} can be regulated to pull down the common mode voltage of the operational amplifier as needed.

[0058] Baseline wander current I_{bl} is also "summed" by the active resistive summer, as shown in Figure 11, to correct baseline wander. Approximately ninety percent (90%) of all system baseline correction can be obtained through the active summer. The remaining baseline residual can be digitally corrected through an equalizer, for

example. As will be appreciated, the Figure 11 topology allows the current sources (I_{b1} and I_{cms}) to each have a fixed output voltage, thus, minimizing current deviation due to finite output resistance.

[0059] The baseline wander correction module 4 preferably corrects for baseline wander using a decision-directed method, such as a discrete integrator. The decision-directed method can be implemented with a known charge pump, where the pump sign (e.g., +1/-1) is determined digitally using an error between the equalized baseline signal (y_k) and a sliced baseline signal ($\hat{y}_k$), as shown in Figure 12. As will be appreciated by those skilled in the art, the expected error value (e.g., $E[e_k]$) is ideally driven to zero. The charge pump is preferably pumped up or down based on the error value. For example, a positive error implies that a negative value should be input into the charge pump. For a negative error, a positive value should be input into the charge pump. The charge pump preferably has at least two current settings to regulate I_{b1}. Of course, a charge pump with many current settings could be used to obtain finer baseline correction control.

[0060] The preferred baseline wander correction performance specifications are further detailed in Table 3, below.

Table 3: Baseline Wander Correction Specification

Parameters	Specifications
Output Dynamic Range	+/-100uA (diff.), (+/-1V/R_1, R_1 = 10kΩ)
Output impedance	High
Integration Factors	2 mV/T, 4 mV/T
Bandwidth	> 100MHz

[0061] A second-order low-pass filter, as shown in Figure 13, is cascaded after the summer to preferably flatten the frequency response out to about 31.25MHz (<1dB). A minimum overall attenuation of 20dB at 125MHz is desirable for the low pass filter. In a sampled system, some aliasing beyond Nyquist frequency (or excess bandwidth) is acceptable, but minimum aliasing is allowed at the sampling frequency. The transmitted data is preferably band-limited to the Nyquist rate.

[0062] Preferred performance characteristics of the low pass filter 5 are further detailed in Table 4, below.

Table 4: LPF Performance Specification

Parameters	Specifications
Input Dynamic Range	+/-1V(diff.)
Output Dynamic Range	+/-1V(diff.)
Input impedance	High, ~10k.

Output impedance	Low
Cutoff frequency	50-60Mhz.
Q (2nd order)	~1
Input impedance	High, ~10k.
Output impedance	Low, <100
DC gain	1

[0063] As an alternative arrangement, a third-order Sallen and Key low pass filter as disclosed in a co-pending application by the same inventor of this application, titled "CALIBRATION CIRCUIT," filed concurrently herewith, and hereby incorporated by reference, could be used as filter 5. Similarly, the calibration circuit disclosed therein could also be used to calibrate the low pass filter 5.

[0064] Analog-to-digital converters are well know in the art. As will be appreciated, the ADC 6 resolution is often determined by system digital processing requirements. In a preferred embodiment, the Viterbi detector 18 requires an effective 7-bit resolution. Residual baseline wander, echoes, and crosstalk increase the dynamic range by about 200-300mV, which increases the required resolution. The reduction in dynamic range due to insertion loss for a 100m cable is approximately 40%. Accordingly, an 8-bit resolution is preferred.

[0065] The preferred ADC performance specifications are further detailed in Table 5, below.

Table 5: ADC Performance Specification

Parameters	Specifications
Resolution	8-bits minimum.
Sampling frequency	8 MHz (125MS)
Source Output Impedance	Low, ~200-400Ω

[0066] As previously discussed, the transmitter 1 preferably includes a current-source I_{tx} that a resistor R_{tx} to generate a transmit signal voltage equal to $I_{tx}R_{tx}$. In one preferred embodiment, a center tap of a transformer is connected to a 2.5 volt supply voltage. If the transmitter 1, for example, transmitting 1 volt or receiving 1 volt, then the possible swing in voltage across resistor R_{tx} is 2 volts. More specifically, the voltage across resistor R_{tx} can vary between 1.5 volts and 3.5 volts. If the power supply voltage source for the circuit is less than the maximum voltage that could be present across resistor R_{tx}, then improper operation or damage can result. It is therefore preferable to provide common-mode shift current control circuitry which will control the amount of current being drawn across resistor R_{tx} so as to keep the composite

voltage signal within an appropriate operating range.

[0067] One embodiment of the present invention is depicted in Figure 15. Common-mode shift current control circuitry is illustrated as common-mode feedback circuitry including an operational amplifier 24 and a pair of field effect transistors (FETs) 25 and 26. The operational amplifier 24 includes a first input terminal, a second input terminal, and an output. The operational amplifier 24 receives a differential voltage signal at the first input terminal, wherein the differential voltage signal is a composite signal which includes components related to the voltage transmitted across the transformer as well as the voltage received across the transformer. A common-mode voltage signal V_{CM} is received at the second input terminal. In the embodiment depicted in Figure 15, V_{CM} = 1.5 volts. The output of the operational amplifier 24 is a common-mode shift current control signal, V_D.

[0068] As will be appreciated by those skilled in the art, when the voltage applied at the first input terminal is different from the voltage applied at the second input terminal, the operational amplifier 24 will sense the inequality. Referring to Figure 16, the internal operation

of the operational amplifier 24 is shown. When the current I_n is greater than the current I_p, the voltages V_p and V_n will not be in balance around V_{CM}. As a result of such an imbalance, V_D will increase. Alternatively, when the current I_p is greater than the current I_n, the voltages V_p and V_n will also not be in balance around V_{CM}. As a result of such an imbalance, V_D will decrease.

[0069] Referring again to Figure 15, the effects of increasing or decreasing V_D are described. By controlling V_D, the common-mode shift current flowing through the resistors R_1 can be controlled. By controlling the common-mode shift current, V_{tx} is thereby also controlled. By being able to control V_{tx}, the situation in which V_{tx} exceeds an operating parameter of the circuit, such as the power supply voltage source, is avoided. Specifically, it is undesirable for the composite voltage signal to exceed the supplied voltage for the circuit. Using the common-mode feedback arrangement depicted in the embodiment of Figure 15, V_{tx} is controlled in relation to the common-mode voltage signal.

$$I_p = I_n \tag{13}$$

$$V_p = V_{cxp} - I_p R_1 \tag{14}$$

$$V_n = V_{cxp} - I_n R_1 \tag{15}$$

[0070] Referring to Equation 13, it should be appreciated that when the common-mode shift currents I_p and I_n are equal, then the voltage drop that occurs over the resistor R_1 is the same for both components of the composite differential signal. See Equations 14 and 15. The differential signal applied to the input of the operational amplifier is therefore of the same magnitude as the differential composite signal.

[0071] It should be appreciated that the common-mode feedback circuit of Figure 15 should work regardless of the resistor value chosen, for example $2R_1$. Because the current drop across the resistors will be the same in both components of the differential signal, the differential voltage signal applied to the input of the active resistive summer will be of the same magnitude as the composite voltage signal. The distinction relates to the output of the summer, and not to the operation or effectiveness of the common-mode shift current control circuitry.

[0072] In another embodiment of the present invention, a constant current source is provided to generate the common-mode shift current control. Referring to Figure 17, constant current sources 27 and 28 are depicted as controlling the common-mode shift currents I_p and I_n.

$$V_{tx2} = V_{tx} - IR_2 \qquad (16)$$

Because I_p and I_n are equal, the magnitude of the differential signal applied to the active resistive summer is therefore shown in Equation 16. As compared with the embodiment in which the common-mode feedback circuit is utilized to control the magnitude of the common-mode shift current, and therefore the magnitude of the composite signal relative to the applied common-mode voltage, in the embodiment utilizing constant current sources, the voltage drop between the composite differential signal and the differential signal applied to the summer will be constant, i.e. IR_2.

[0073] In another embodiment of the present invention, the common-mode shift current control circuitry includes a resistor divider. Through the use of a resistor divider, the voltage signal applied to the summer is a

proportionately reduced signal as compared with the composite voltage signal.

$$V_{tx3} = V_{tx}R_2/(R_1 + R_2)$$

It should be appreciated by those skilled in the art that, by selecting appropriate values for resistors R_1 and R_2, the magnitude of the common-mode shift current and the composite signal can be controlled so that the magnitude of the composite signal does not exceed an operating parameter of the communications circuit, such as the power supply voltage source.

[0074] Thus, a transmit canceller including an active resistive summer has been described. Such an active resistive summer has not heretofore been developed for applications such as canceling signals in gigabit channels. Correcting baseline wander through such an active resistive summer has also been described herein. Controlling common-mode shift current has also been described herein.

[0075] While the present invention has been described with respect to what is presently considered to be the preferred embodiments, it will be understood that the

invention is not limited to the disclosed embodiments. To the contrary, the invention covers various modifications and equivalent arrangements included within the spirit and scope of the appended claims. The scope of the following claims is to be accorded the broadest interpretation so as to encompass all such modifications and equivalent structures and functions.

[0076] For example, while preferred circuit configurations and component values have been described, it will be understood that modifications could be made without deviating from the inventive structures. For example, values for the feedback and input resistors R_f and R_1 could be changed to obtain higher or lower gains. Also, an active resistive summer could be constructed to sum only the measured signal V_{tx} and the replica signal V_{txr} (or a high-pass version of the replica), for example. Additionally, while the communication channel has been described with respect to a twisted pair of cables, the invention may also be practiced with other communication channels such as optical and wireless channels. Moreover, this invention should not be limited to gigabit transmission rates and can be practiced at any transmission rate requiring the signal processing characteristics of the

invention. Of course, these and other such modifications are covered by the present invention.